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Computer Compensation for Cable Signal Degradations



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COMPUTER COMPENSATION FOR CABLE SIGNAL DEGRADATIONS

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December 1987

ABSTRACT

This paper describes two techniques for computing software cable compensation filters. These filters are used in correcting waveforms recorded from diagnostics on pulsed power accelerators. Applicable topics in continuous and discrete linear systems theory are reviewed. The first technique for computing a compensation function consists of recursively solving a discrete time domain convolution equation using measured undegraded and cable degraded pulses. The second technique computes the compensation function in the frequency domain using an analytical model of the cable frequency response and a constrained inverse filter. Detailed procedures are described for computing cable compensation filters using an interactive data manipulation and hardware control program.

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1. Summary

The data recording facilities for most pulsed power accelerators at Sandia are located such that a few hundred feet of coaxial cable are required to route the signals from the diagnostics to the recording hardware. This is due to the large size of the accelerators, the need to protect the recording hardware from EMP and radiation, and the convenience of having all the equipment in a single shielded room. These data transmission cables act as low pass filters and attenuate high-frequency components of the signals they transmit. The degree of attenuation depends on both the type and length of cable used. The high-frequency components can be effectively restored by passing the signal through a special filter.

Such a filter may be implemented by either a hardware equalizer¹ or, if the waveform is digitized, by a software digital filter. This report describes techniques for creating software compensation filters using a computer. Two methods are presented. The first one consists of solving the discrete time domain convolution equation recursively. The second method consists of analytically modeling the cable frequency response and then performing a constrained inversion. Both techniques are generally known as deconvolution. Compensation filters are computed using a specially developed interactive data manipulation and hardware control program called IDR. Different versions of this program have been implemented in various facilities. Detailed procedures for creating the compensation filters are described for the software and hardware in the data acquisition facility for the Particle Beam Fusion Accelerator II (PBFA II).

2. Cable Compensation Theory

This section describes some basic theoretical aspects of restoring degraded signals to their original form. First, elementary relations in linear system theory are given.² The limitations imposed by noise are discussed. Finally, both the time and frequency domain techniques used to restore signals are described.

2.1 Linear System Equations

2.1.1 Fourier Transform

The frequency-space distribution $F(\omega)$ of a time-varying signal $f(t)$ may be computed by taking the Fourier transform

$$F(\omega) = \int_{-\infty}^{\infty} f(t) e^{-j\omega t} dt \quad (1)$$

$F(\omega)$ is generally complex.

The original signal $f(t)$ may be recovered from $F(\omega)$ by taking the inverse Fourier Transform

$$f(t) = (1/2\pi) \int_{-\infty}^{\infty} F(\omega) e^{j\omega t} d\omega \quad (2)$$

2.1.2 Delta Function

The Dirac delta function is defined as

$$\begin{aligned} \delta(x-x_0) &= \infty & x &= x_0 \\ &= 0 & x &\neq x_0 \end{aligned} \quad (3)$$

$$\int_{-\infty}^{\infty} \delta(x-x_0) dx = 1 \quad (4)$$

The delta function Fourier transform pair is

$$\Delta(\omega, x_0) = \text{FT} [\delta(x-x_0)] = e^{-j\omega x_0} \quad (5)$$

$$\Delta(\omega, 0) = 1 \quad (6)$$

$$\text{FT}^{-1}[1] = \delta(x) \quad (7)$$

where $FT[\]$ is the Fourier transform of the function in the brackets, and $FT^{-1}[\]$ is the inverse Fourier transform.

2.1.3 Time-Invariant Systems

If an input signal $f(t)$ is passed through a linear time-invariant system with impulse response $h(t)$, the output $g(t)$ may be computed from the convolution integral

$$h(t) = \int_{-\infty}^{\infty} f(\tau) h(t-\tau) d\tau = \int_{-\infty}^{\infty} f(t-\tau) h(\tau) d\tau \quad (8)$$

The function $h(t)$ is the output of the system when the input is $\delta(t)$:

$$h(t) = \int_{-\infty}^{\infty} \delta(\tau) h(t-\tau) d\tau \quad (9)$$

The convolution process will be denoted by the symbol "*":

$$g(t) = f(t) * h(t) = h(t) * f(t)$$

Convolution in the time domain is equivalent to multiplication in the frequency domain:

$$\begin{aligned} G(\omega) &= FT[g(t)] = FT[f(t)*h(t)] \\ &= F(\omega) H(\omega) \end{aligned} \quad (10)$$

2.1.4 Discrete Approximations

The zero-order discrete approximation to the convolution integral is

$$g_n = T \sum f_i h_{n-i+1} \quad \text{for } n = 1, 2, \dots, N \quad (11)$$

where T is the sampling interval for the sequences of digitized data f , g , and h ³:

$$g_n = g(nT)$$

$$f_i = f(iT)$$

N is the number of points in f , g , and h .

Equation 11 may be solved recursively for the impulse response h if both f and g are known. This process is called deconvolution. The set of equations defined by Eq. 11 is:

$$\begin{aligned} g_1 &= T f_1 h_1 \\ g_2 &= T(f_1 h_2 + f_2 h_1) \\ g_3 &= T(f_1 h_3 + f_2 h_2 + f_3 h_1) \\ &\vdots \end{aligned}$$

Solving these consecutively yields the sequence h :

$$\begin{aligned} h_1 &= g_1 / (T f_1) \\ h_2 &= (1/f_1) ((g_2/T) - f_2 h_1) \\ h_3 &= (1/f_1) ((g_3/T) - f_3 h_1 - f_2 h_2) \\ &\vdots \\ h_n &= (1/f_1) ((g_n/T) - \sum_{i=2}^n f_i h_{n-i+1}) \end{aligned} \quad (12)$$

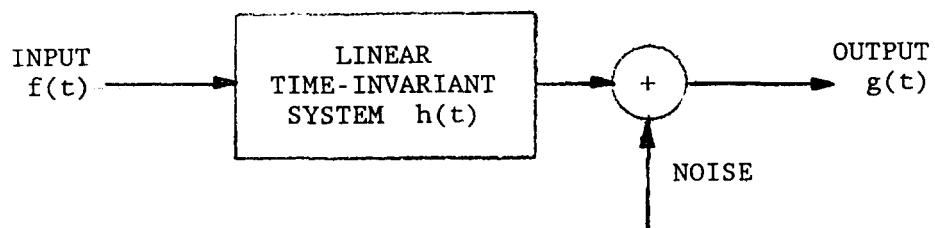


Figure 1. Block diagram of the model of the cable system.

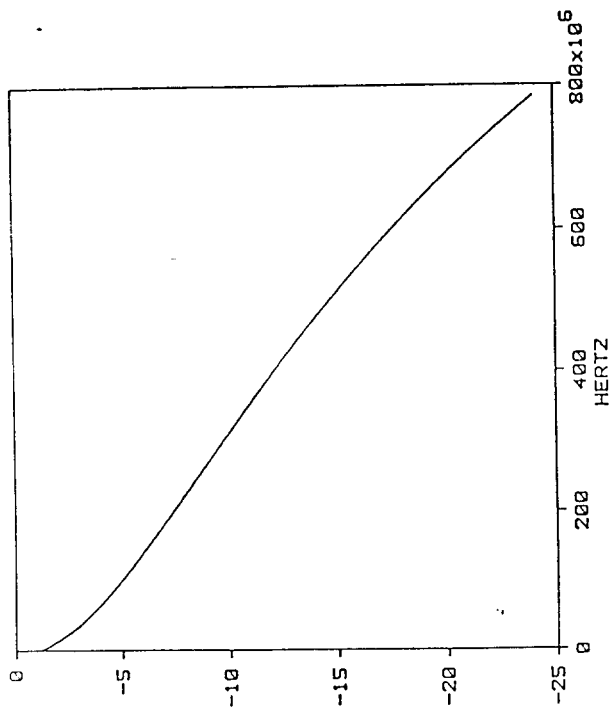


Figure 2a. Typical cable frequency response.

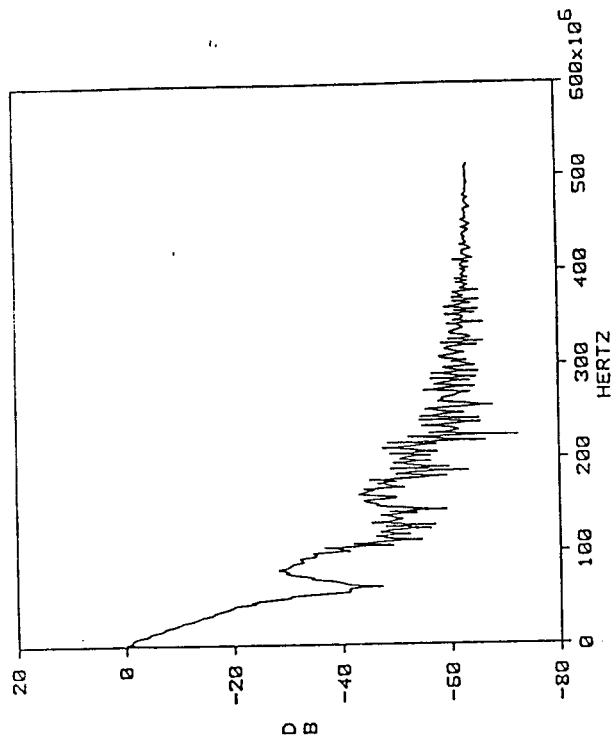


Figure 2b. Typical signal frequency spectrum.

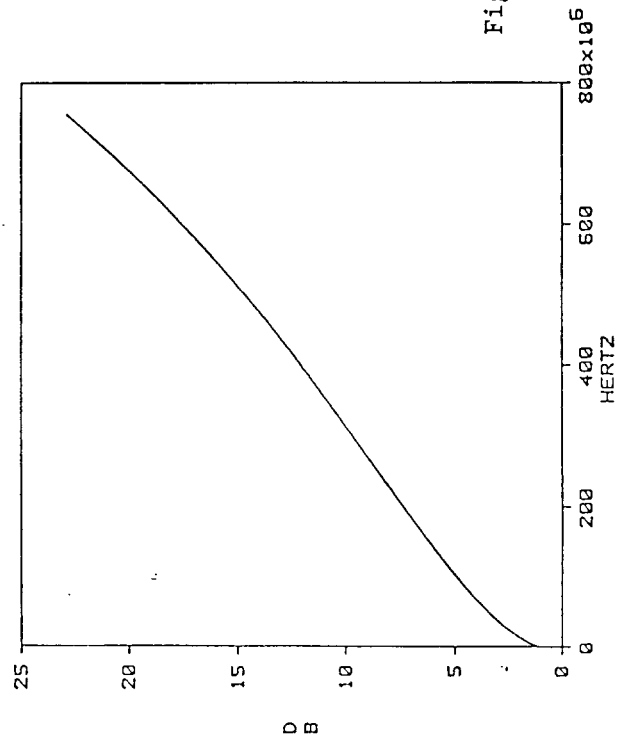


Figure 2c. Inverse of cable frequency response.

2.2 Signal Restoration

The model of a linear time-invariant system will be used to characterize the effects of a cable on an input signal. Let the cable have an impulse response $h(t)$ and corresponding Fourier transform $H(\omega)$. If the measured signal $g(t)$ were noise free, the original signal could be recovered by passing $g(t)$ through an "inverse" filter with response

$$r(t) = \text{FT}^{-1}[1/H(\omega)] \quad (13)$$

assuming $H(\omega)$ has no zeros.

The symbol " $\hat{}$ " will be used to denote the restored signal.

$$\begin{aligned} \hat{F}(\omega) &= G(\omega)/H(\omega) \\ &= F(\omega) H(\omega) (1/H(\omega)) \\ &= F(\omega) \end{aligned} \quad (14)$$

However, if $g(t)$ does contain noise added at the output, as shown in the system block diagram in Fig. 1, we have

$$\begin{aligned} g(t) &= f(t) * h(t) + e(t) \\ G(\omega) &= F(\omega) H(\omega) + E(\omega) \\ F(\omega) &= F(\omega) + E(\omega)/H(\omega) \end{aligned} \quad (15)$$

The following relationships are generally true for large ω :

$$\begin{aligned} |E(\omega)| &= \text{constant} \\ |F(\omega)| &= 0 \\ |H(\omega)| &= 0 \\ F(\omega) &= E(\omega)/H(\omega) \end{aligned} \quad (16)$$

Thus high frequency noise is strongly amplified and indeed is unbounded. These relations are illustrated graphically in Fig. 2 a,b,c. The amplified noise typically totally masks the signal in restorations using this approach. Thus for a useful restoration, the simple inverse filter must be constrained in some manner.

2.3 Constrained Restoration Techniques

There are many techniques for constraining the inverse filter restoration technique. The time-domain method we will describe was developed by Thane Hendricks at EG&G Las Vegas.⁴ The frequency-domain method involves a heuristic modification to the constrained least squares method of deconvolution.

2.3.1 Time Domain Cable Compensation

The time domain deconvolution method is based on equation (12). The solution for each term of the compensating impulse response contains the term $1/f_1$. Thus f_1 cannot be 0. Also, since the solution is recursive, i.e., it uses past values of h_i to compute the next one, it is potentially unstable. The instability occurs when f_1 is too small. When this is the case the filter coefficients will increase exponentially and "blow up". This has been observed to happen many times in practice. The method developed by Hendricks is to relate the size of f_1 to the maximum frequency domain gain in the compensation function. This method is referred to as "truncating the toe of the step response". The steps in the algorithm are outlined below.

1. Characterize the cable by first recording as close an approximation to an ideal unit step, $u(t)$, as possible directly at the input of the waveform recorder. The recorded approximation to a unit step will be called the undegraded signal and will be denoted as $u'(t)$. Then apply the same step to the input of the cable under test. Record the cable step response, $s(t)$. This signal will be called the degraded pulse. A theoretical step response is shown in Fig. 3a. Fig. 3b shows the corresponding impulse response.

2. Truncate the leading edge or "toe" of $s(t)$ to form

$$\begin{aligned} s'(t) &= u(t) s(t+t') & t \geq 0 \\ &= 0 & t < 0 \end{aligned} \quad (17)$$

3. Use the deconvolution equation (12) to calculate the restoring function impulse response r_n .

$$r_1 = u'_1 / (Ts'_1)$$

$$r_n = (1/s'_1) ((u'_n/T) - \sum_{i=2}^n s'_i r_{n-i+1}) \quad \text{for } n=2,3,\dots,N \quad (18)$$

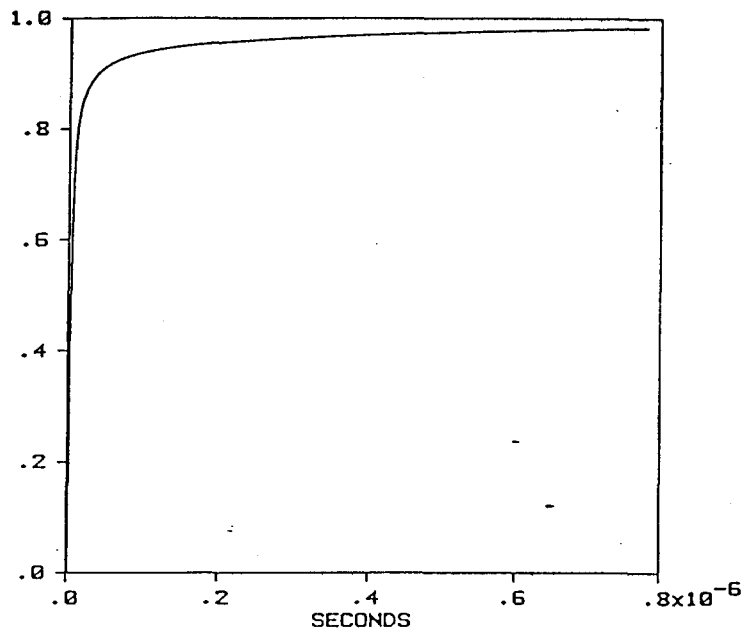


Figure 3a. Step response of a typical cable run.

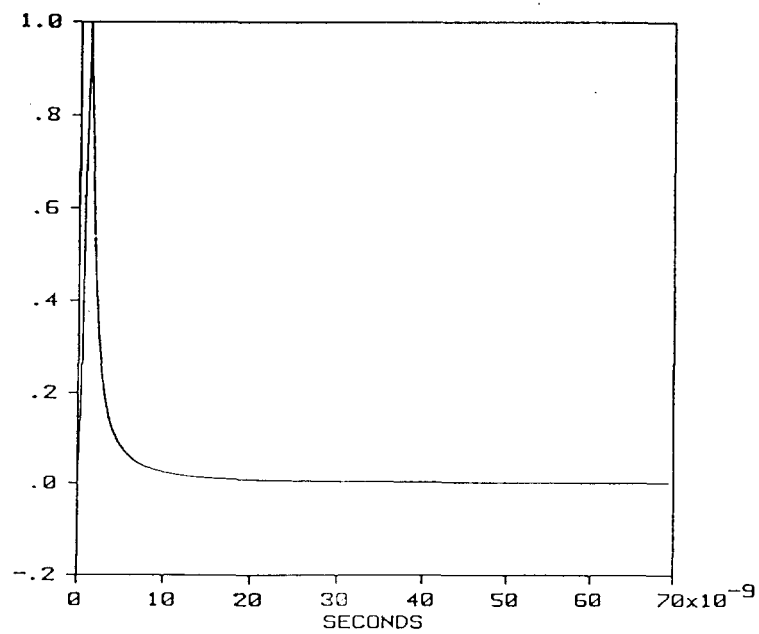


Figure 3b. Impulse response of a typical cable run.

4. The measured signal waveform is then directly convolved with the filter coefficients r_n in the time domain using equation (11).

$$f_n = \sum_{i=1}^n g_i r_{n-1+i} \quad (19)$$

This operation requires $N/2$ floating point operations. To speed up the computation for large N , the computation may be done in the frequency domain using Fast Fourier Transform (FFT) methods. Both the data and filter waveform must be padded with zeros to twice their original length to eliminate wraparound effects associated with frequency domain convolution.³

There are two reasons for truncating the toe of the step response function $s(t)$. First it insures that s_1 is nonzero. The second reason can be seen by examining its frequency domain implications. The impulse response of a system is the time derivative of its step response

$$h(t) = ds(t)/dt$$

The impulse response corresponding to the toe truncated step is

$$\begin{aligned} h'(t) &= ds'(t)/dt \\ &= d[u(t) s(t+t')] / dt \\ &= s(t') \delta(t) + u(t) h(t+t') \end{aligned} \quad (20)$$

where

$u(t)$ is the ideal unit step.

In the frequency domain we have

$$\begin{aligned} H'(\omega) &= s(t') + \text{FT}[u(t) h(t+t')] \\ |H'(\omega)| &= s(t') + |H(\omega)| \end{aligned} \quad (21)$$

This function and its inverse are plotted in Fig. 4. Note that the frequency response of the constrained inverse levels off rather than continuing to rise. The truncation fraction is the inverse of the maximum amplification. In the example shown in Fig. 4 the truncation fraction is .1, and the maximum amplification in the frequency domain is 20 dB. In practice the undegraded step signal used for $u(t)$ is not a perfect step, and the maximum gain is smaller than the inverse of the truncation fraction.

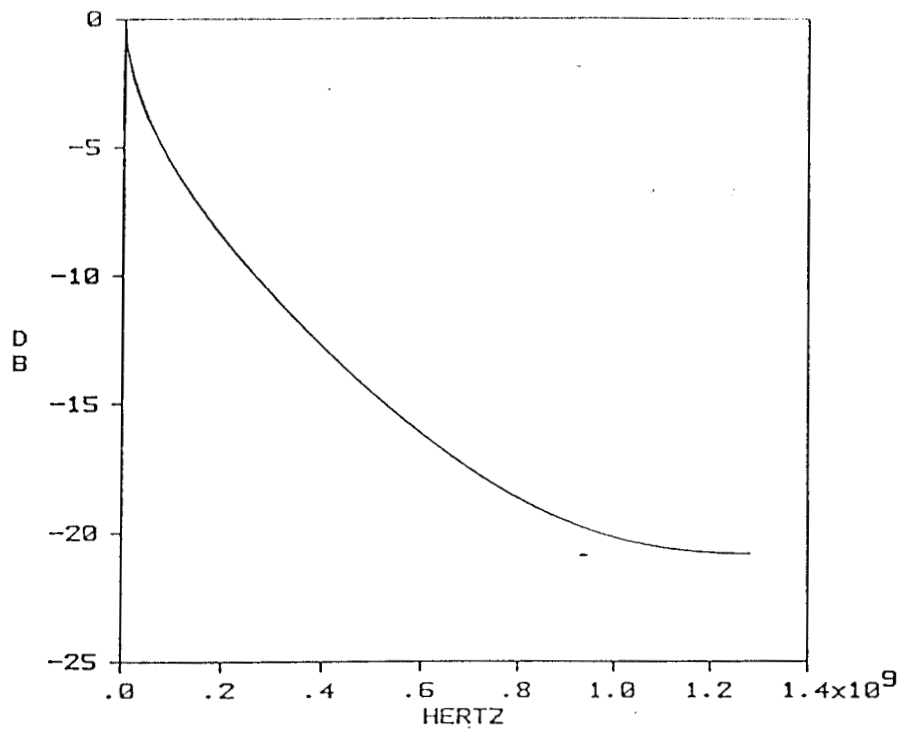


Figure 4a. Frequency spectrum of a step response with a toe-truncation fraction of .1.

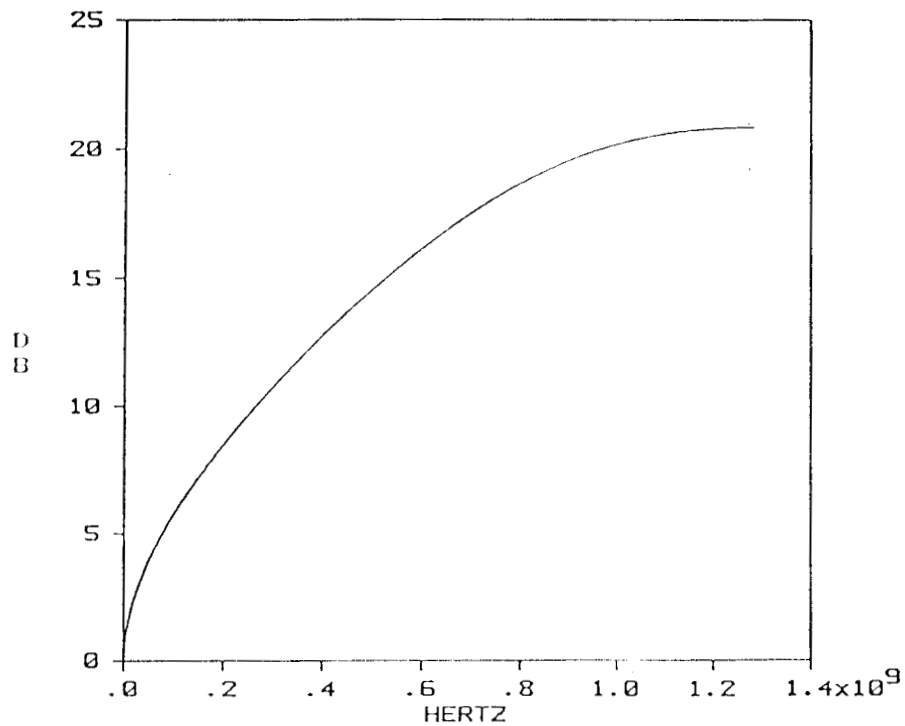


Figure 4b. Inverse of the frequency response of the truncated cable step.

2.3.2 Frequency Domain Analytical Method

We showed in equation (16) that simply inverting the cable frequency response to restore the signal will not work in the presence of noise. We can modify equation (14) to eliminate excessive noise amplification.

$$\hat{F}(\omega) = \bar{H}(\omega) G(\omega) / [\bar{H}(\omega) H(\omega) + C(\omega)] \quad (22)$$

where $C(\omega)$ is a constraint function. The overbar indicates the complex conjugate of the quantity. The properties of $C(\omega)$ are that it is small for low frequencies and large for high frequencies. An appropriate form is

$$C(\omega) = \gamma \omega^p \quad (23)$$

where γ and p are constants to be determined. In the deconvolution method known as Constrained Least Squares the constants γ and p are often chosen to minimize the second derivative of the chosen signal. In this case $p=4$ and γ must be found iteratively.⁵ In our method of computing cable compensation filters, p is chosen subjectively; γ is then chosen to achieve a specified 3 dB compensated bandwidth.

It is possible to estimate the impulse response of the cable by time domain deconvolution of a measured degraded and undegraded step pulse using equation (12). This again requires careful selection of a truncation level in the recorded undegraded pulse. Then the frequency response can be found by taking the Fourier transform. An alternate method is to use an analytical model of the cable's frequency response. An appropriate model is:⁶

$$H(\omega) = e^{-\alpha\sqrt{j\omega}} \quad (24)$$

The attenuation constant α may either be measured or determined from manufacturers' tables and curves. Since there is only one constant, it may be determined from a single attenuation measurement; or a least squares fit may be done using multiple measurements.

There are two digital signal processing problems associated with this method. First the assumed function rolls off rapidly at low frequencies, but the rolloff is much slower at high frequencies. In many cases the response may be down by less than 20 dB at the Nyquist frequency. This results in excessive aliasing. We have found that windowing the frequency domain equation with a "half-cosine" function eliminates the aliasing and does not degrade the restoration. The window function used is

$$w(\omega_i) = .5[1 + \cos(\omega_i\pi / ((N-1)T))] \quad (25)$$

where the ω_i are the frequencies implicit in the FFT, i.e.,

$$\omega_i = (i-1) 2\pi / (NT) \quad (26)$$

The second digital signal processing problem associated with this method is that simply using a windowed equation (20) does not produce a causal cable impulse response, i.e., the impulse response starts before $t=0$. Due to the periodic nature of the Discrete Fourier Transform (DFT), this manifests itself by the leading edge of the impulse response occurring at the very end of the time domain record. We have solved this problem by introducing an arbitrary time delay, Δt , in both the cable and restoration function frequency response. Thus the net equation for the cable frequency response is

$$H(\omega_i) = w(\omega_i) e^{\alpha\sqrt{j}\omega_i} e^{-j\omega_i\Delta t} \quad (27)$$

and the net equation for the restoration function frequency response is

$$R(\omega_i) = e^{-j\omega_i 2\Delta t} \bar{H}(\omega_i) / (\bar{H}(\omega_i) H(\omega_i) + \gamma \omega_i^p) \quad (28)$$

The value of p is typically set to 2 or 4. The impulse response of the restoration function may be found by taking the inverse Fourier transform of $R(\omega)$. The degraded signal is then restored by either time- or frequency-domain convolution with the restoration function.

The amplitude of the cable frequency response rolls off as $e^{-\alpha\sqrt{j}\omega}$ or $e^{-\alpha(j\omega)^{.5}}$ in the analytical model. Manufacturer's curves of cable frequency response indicate that the rolloff exponent is typically higher than .5. For example, the exponent for RG-214 is .565; for RG-331 it is .55. During testing we found that the shape of the cable step response was strongly dependent on the exponent used. Fig. 5 shows the computed step responses for exponents of .5 and .56 for the total PBFA II water section cable run. Thus in practice the equation (27) should be replaced by the equation below for the cable frequency response,

$$H(\omega_i) = w(\omega_i) e^{\alpha(-j\omega_i)^\beta} e^{-j\omega_i\Delta t} \quad (29)$$

and β is chosen to match the actual cable rolloff exponent.

Details of computing cable compensation filters and actual results are given in the next sections.

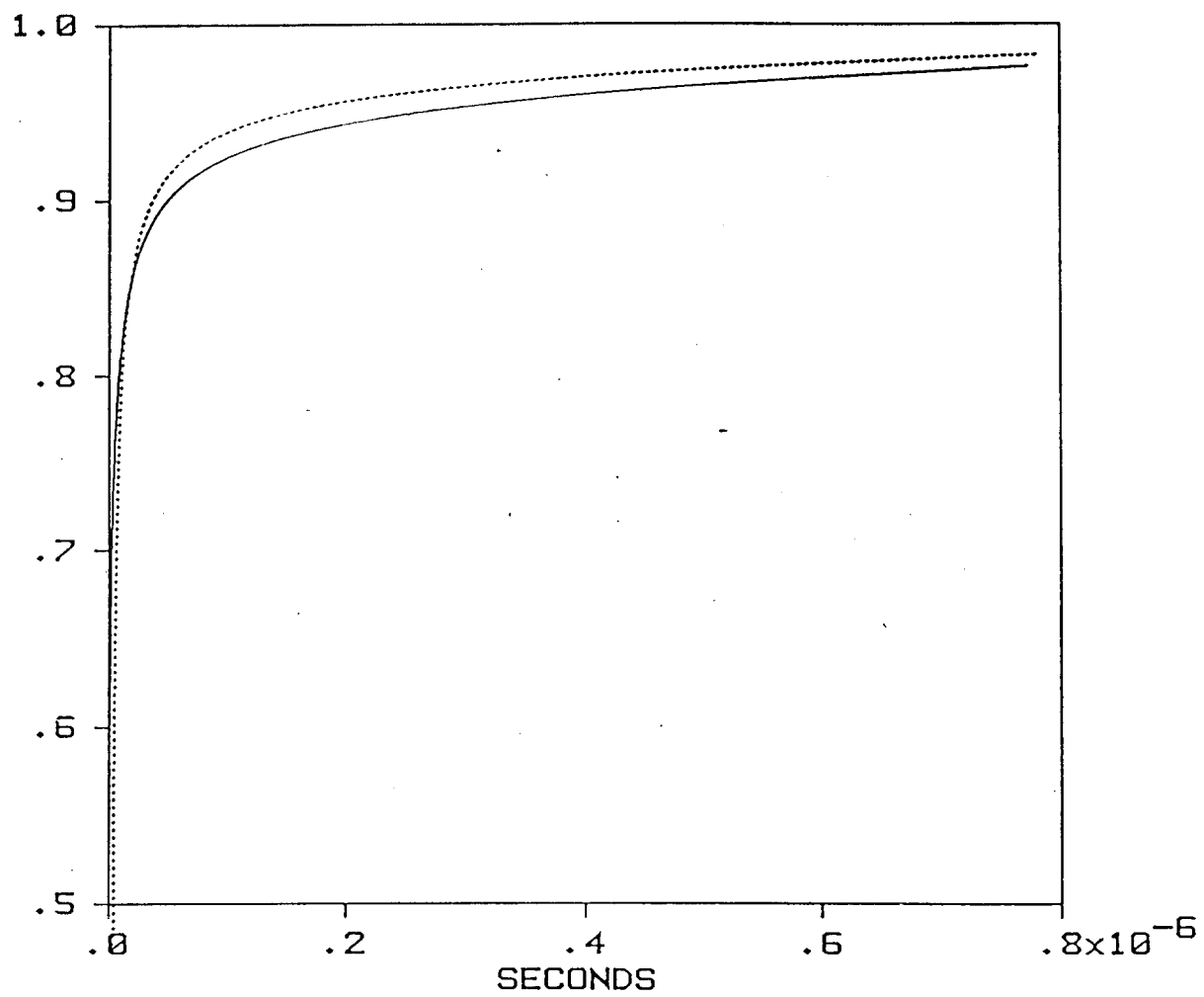


Figure 5. Analytical cable step response for a cable rolloff coefficient of .5 (solid line) and .56 (dotted line).

3. Cable Compensation in Practice

Most cable runs consist of segments of different types of cables. It is most convenient to measure all the segments in a cable run together as a single system. If there are many identical runs, the compensation filter only needs to be computed for one of them. A different cable compensation filter must usually be designed for different cable runs. We have not developed firm guidelines for how much extra cable of a given type can be added to an existing run before a new compensator is required. As will be noted in the procedures below, there is some leeway or uncertainty in designing the compensation filters. In the time-domain method, the uncertainty is in locating the best toe truncation level. In the frequency-domain method, the uncertainty is in selecting the 3 dB down frequency, the rolloff exponent, and the cable attenuation exponent. One possible criteria is to note the frequency at which the compensation filter has its maximum gain. For PBFA II using Tektronix 7912AD waveform recorders, this is typically 200 MHz. If adding a segment of cable changes the attenuation at 200 MHz by more than 5%, a new compensator should be computed. The cable attenuation program, CAT, can be used to compute the frequency response of different cable runs.

The PBFA II Data Acquisition System (DAS) has Tektronix 7912AD and LeCroy 6880 waveform recorders. Individual cable compensation filters are required for each 7912AD sweep speed and for the 6880's. There are two reasons for this. First, for a given compensator, the filter coefficients vary rapidly for the first 10 or 20 elements. This variation makes the function unsuitable for interpolation. Second, the effective width of the sampling pulse limits the bandwidth of the recorder. The 7912AD's have an effective sampling pulse width of at least 3 or 4 times the sampling time. This is due to the finite width of the scan converter write and read beams. Thus the actual bandwidth decreases for increasing sweep time.

The current version of the LeCroy 6880's have two problems that limit cable compensation effectiveness. First they introduce signal dependent colored noise onto pulse type signals. The baseline typically has white noise with a standard deviation of .6 counts. But the noise on top of a pulse with a 130 count amplitude has a noise standard deviation of 3.0 counts. This signal dependent noise has frequency components at multiples of 42 MHz. This is the basic clock frequency of each Charge Coupled Device (CCD) used to store the analog waveform. The 6880 has 32 CCD's. The second problem in computing cable compensations is that the CCD's are continuously storing samples before the trigger signal. Thus the CCD's are not synchronized to the signal from record to record. Due to different transmission paths in the commutators, the 6880 has 32 different step responses depending on which CCD stored the first sample after the trigger pulse. These two problems combined with the inherently unstable nature of the time-domain deconvolution equation preclude computing full 250 MHz

bandwidth cable compensation filters on the 6880's as is done on the 7912AD's. We have developed the following solutions to this problem:

- Low-pass filtering both the degraded and undegraded pulses to 75 MHz before performing the time domain deconvolution. This produces a well-defined compensation function, but limits the system bandwidth to about 100 MHz.
- Using hardware cable equalizers. These work well. The disadvantages are that they attenuate the signal and that a special equalizer must be purchased for each different cable run.
- Using the analytic frequency domain method .
- Averaging multiple undegraded and degraded pulses to reduce noise and average out differences in pulse response. This method works well also. The averaging method reduces the risetime of the undegraded pulse slightly, $\approx .2$ ns, which reduces the final compensated bandwidth to about 190 MHz. This is the method currently used on the PBFA II DAS. Either 16 or 32 pulses are averaged for both the degraded and undegraded signals. The IDR COMPARE command is used to compute a time shift by which each pulse is adjusted before being added to the running average.

The software cable compensation introduces a time delay into the compensated signal. The time delay is approximately equal to the time between the start of the filter and the first positive peak. For single pulse compensators computed on 7912AD's, the time delay is only about 1 sample time. The analytical frequency domain method and the time domain method using averaged pulses introduce 2 to 7 ns of delay. The delay is compensated for by setting the t-zero time of the compensation filter to minus the time from the start of the signal to its peak. The software that performs the cable compensation filtering then uses this value to time shift the compensated waveforms back to their proper time.

The next two sections describes how to actually compute time domain cable compensation filters using the interactive data reduction and hardware control program IDR. This program supports all the commands necessary to record the required signals, perform the manipulations on the data, and store the compensation filters in the special file. The instructions below are specifically for the version of IDR and the system hardware in the PBFA II DAS. The detailed instructions are given for computing time domain compensation filters using the 7912AD recorders. Differences in the procedure for using the 6880 recorders are outlined later. Reference is made to other computer programs in the system. These are: CAT, the cable attenuation computation program; AQD, the automatic data acquisition program; and ACL, the automatic waveform recorder calibration program. Detailed instructions for using these programs are available in on-line help files on the DAS computer.

3.1 Time Domain Cable Compensations Using IDR

The time domain method requires that the undegraded and degraded step pulses be digitized. The hardware setups to measure these pulses are shown in Fig. 6. The pulse generator should have a risetime at least twice as fast as the recorder. The pulse amplitude should be about half of full scale. The pulse width should be wide enough so that the degraded pulse has essentially reached its final value. This condition cannot be achieved for fast sweep speeds and long cables on 7912AD's. First, the step output is measured at the digitizer, and this signal is saved. The pulse is then sent down the cable to be characterized and digitized. Care must be taken to insure that the output pulse amplitude does not change when the pulse generator is moved.

A Tektronix PG 506 calibration pulser works well for computing cable compensation filters. The positive slope fast transition is normally used. The only problem with this signal is that the transition is from a negative level to ground. In order to use this pulse, it must be shifted using the ADD command so that the transition starts at zero. Extreme care must be taken to insure that the amplitude adjustment is not changed when the pulser is moved from the recorder to the end of the cable. One solution to this potential problem is to set the amplitude at full scale and use fixed attenuators to get the proper level. Another acceptable pulse generator is the Picosecond Pulse Labs model 2500. The 250 ns width is sometimes too short for long cable runs to reach equilibrium. The procedure for computing cable compensation filters requires using the interactive data reduction computer program IDR. It will be assumed that the reader is familiar with IDR commands and operations. Typical IDR command sequences will be shown below each major step.

1. Record the output pulse from the pulser by with as short a cable as possible from the pulser to a 7912. The pulse width should extend to the end of the record for the sweep speed being used. The pulse amplitude into the 7912 should be set to the maximum, which is about 1 volt. The vertical amplifier should be set to .2 v/div, DC coupling, and full bandwidth. The mainframe intensity and focus must be set to the values stored in the time base calibration record for the unit and sweep speed being used. The IDR commands MSE, TSE, and VSE may be used to set the various parameters.

The TV and Local modes of the 7912 may be used to set proper trace vertical and horizontal positions. Note whenever Local mode is used, all 7912 mainframe and plug-in settings that use knobs (as opposed to push buttons) are reset to the knob value. These are vertical position, horizontal position, trigger level, trace intensity, and focus. It is very important that the intensity and focus settings be set to the proper value using the MSE command after exiting LOCAL mode. The proper mainframe and intensity and focus settings may be determined by running the program ACL and printing the time base calibration record for the unit and sweep speed being used. For example, to print this information for unit 2 at 20 ns/div enter the ACL command:

PRI,2,TB,20.

The test setup for doing cable compensations is shown in Fig. 6. Both undegraded and degraded pulses must be recorded for each sweep speed desired. All of the pulses recorded and created during a compensation run should be stored in a shot data file. The program AQD must be run with a dummy header to create the data file and store the shot number, date, time and label in the shot status file. To do this create a header using one of the text editors. The header only needs to have the label line. Then run AQD with the proper header name and shot number. Next, run IDR with the proper machine name and shot number. Typically the machine name DASCHR is used. The GO form of the program run command must be used since hardware control is required.

GO, IDR, DASCHR, 555 - define output data file

Cable compensation functions are stored as impulse response waveforms in a scratch waveform file called CABCF. All programs that perform cable compensations, AQD, IDR, and ACL, search this file for the requested compensator. The commands for setting up the 7912AD and recording the pulses are shown below with explanatory comments. The comments are not part of the command.

SEL 1	- select unit 1, a 7912AD
SWI,1,1	- set coax switch
PAS,1,0	- turn off prog atten
TV	
LOC	- adjust settings manually while continuously triggering the 7912 and the pulse generator
REMOTE	
MSE MAI 325;FOC 15	- example intensity and focus
MSE MOD DIG	- place unit back in digital mode
VSE INP GND	
TSE SRC LIN	- use line trig for baseline
DIG	- digitize a baseline
ACQ	
NOR,A	
BAS,A	- save baseline for later cal
VSE INP A	
TSE SRC INT	- external trigger may also be used
DIG	- digitize the pulse
ACQ	- read the raw data
NOR,A	- average to center of trace
CAL,A	- remove baseline; convert to volts; correct for time base.
LAB,A,UNDGP20	- last 2 characters are sweep speed
PLOT,A	- see Fig. 7
DWR,A	- save pulse in output file

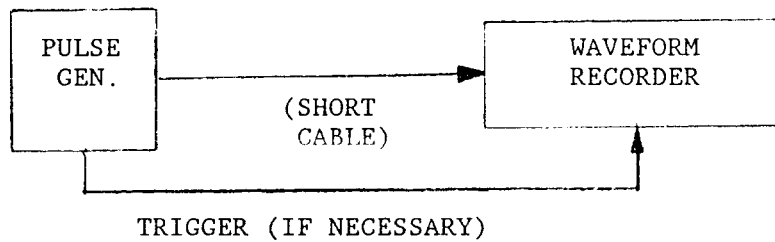


Figure 6a. Test equipment configuration for measuring the undegraded pulse.

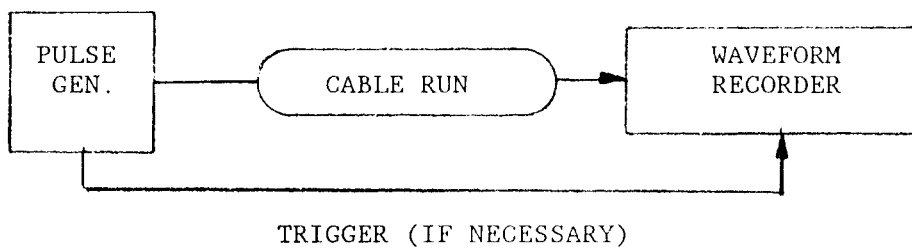


Figure 6b. Test equipment configuration for measuring the degraded pulse.

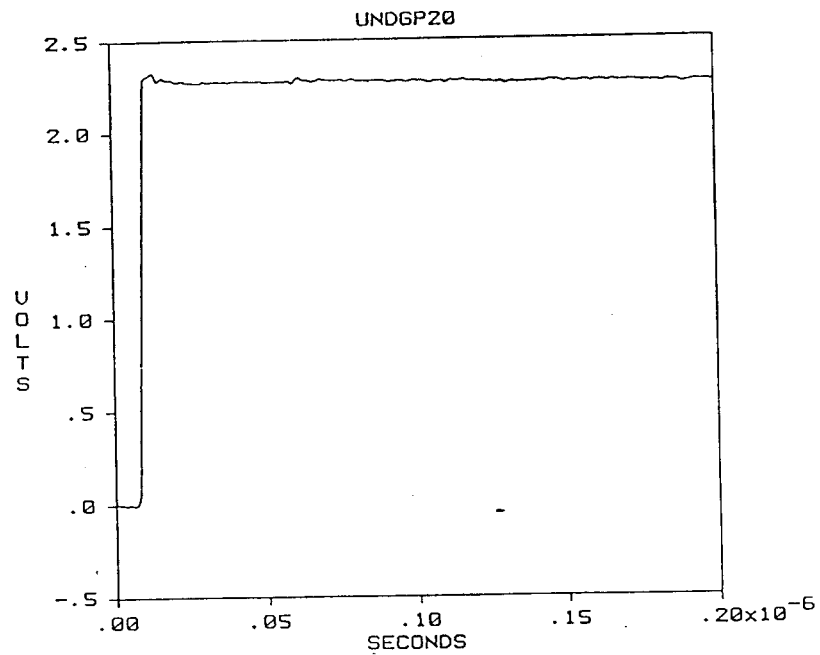


Figure 7. Undegraded pulse measured on a Tektronix 7912AD at 20 ns/div.

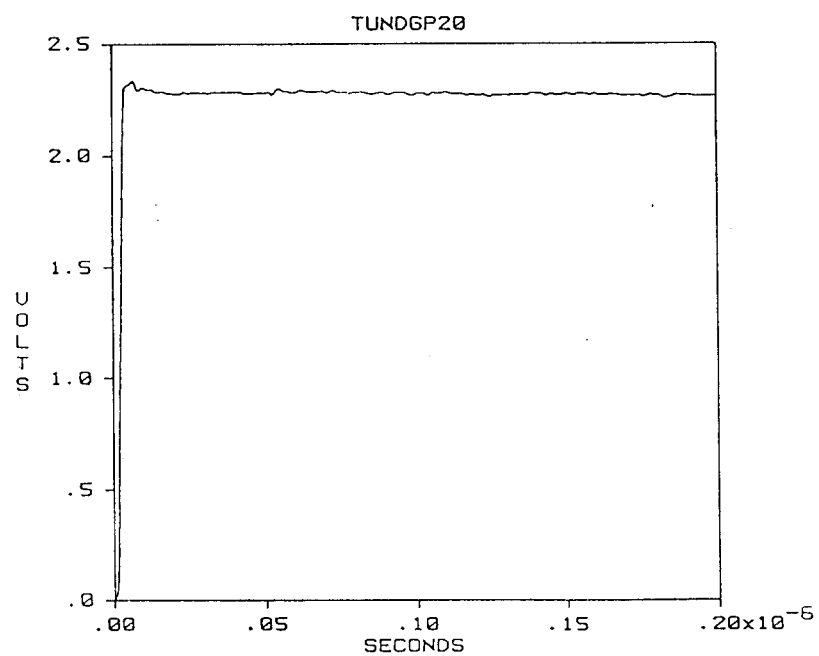


Figure 8. Undegraded pulse shifted so it starts at 0,0.

2. In order to properly compute a cable compensation function, the leading baseline on the pulse must be eliminated so that the leading edge of the pulse starts at 0,0. Also, the input signal must look very much like a step pulse. It must rise to a level and stay there, i.e., it must not return to zero within the record. This requirement sometimes cannot be met at slow sweep speeds with a short pulse. There is an IDR command file called TRUNC which will truncate the leading edge and repeat a given sample out to sample 512. The run command for TRUNC is:

```
RUN,TRUNC,array,tshift,repeat_sample
    array - array register containing the pulse. A,B,C etc

    tshift - time in seconds to desired point on the leading
              edge to truncate to. Tshift must be negative.
    repeat_sample - sample number whose value is repeated
                    out to sample 512. Since the repeating
                    is not done until after the truncation,
                    the number of samples truncated must be
                    subtracted from any repeat_value
                    determined by examining the untruncated
                    pulse.
```

The crosshair cursors are useful for determining the proper value for tshift. The cursors may also be used to determine repeat_sample. The value chosen for repeat_sample is not critical; and it is most often selected by simply eyeballing the data. The procedure for properly shifting the undegraded pulse is shown below.

```
CUR                                - place cursors at point where
X=.24564 E-7                        pulse breaks from 0
Y=.10000 E-2                        - start of pulse
RUN,TRUNC,A,-24.56E-9,480          - shift array left and repeat
                                   sample 480 out to 512.
PLOT,A                             - see Fig. 8
PRINT A
```

The array printout should be examined to determine if the first sample is nearly zero. Following samples should increase monotonically along the rise of the pulse. If there is more than one zero sample, the pulse must be truncated more. If the truncation went too far, i.e. the first sample is not zero, the pulse must be read back from the disc file and truncated with less of a time shift. If the pulse rises monotonically but slowly, truncate it so the second sample is at least 5% of the maximum.

```
CHANGE A 6 0.                     - force first point to be zero
LAB,A,TUNDP20                      - label the signal
DWR,A                              - save in disc file.
```

3. Take the pulser out to the end of the cable for which compensation is desired. Do not disturb the amplitude setting. Record the degraded pulse and save it in the disc file. Use the same

commands shown in step 1. A different label must be used for the degraded waveform. It is recommended that the letters "DGR" replace the "UND" in the label. The degraded pulse should look like that shown in Fig. 9.

4. The key step in computing the compensation function is to truncate the leading edge of the degraded step pulse at some appropriate level. The lower the level, the more high-frequency gain will be applied to the compensated signals. If too low a level is chosen, the compensated pulse will have excessive ringing. If the level is very low, the compensation function computed in the deconvolution described below will "blow up" and may cause the program to abort due to a floating point overflow error. Choosing a level is done iteratively. The criteria for selecting the final value is described later. The degraded signal is truncated using the TRUNC command file as shown in step 2. A good starting truncation level is about 25% of the maximum pulse level. The optimum truncation level will decrease with longer cables and increase with slower sweep speeds. At very low sweep speeds a problem may arise where there is no signal sample at or near the desired truncation level. In this case the IDR CHANGE command should be used to create a sample at the desired level. There is no discernable error introduced in doing this. Another help in choosing a truncation level is to print the first 50 samples of the degraded signal before and after truncation using the PRINT command. A typical truncated pulse is shown in Fig. 10. The truncated degraded pulse must rise to a level and stay there. Like the input pulse it must not fall. Again the TRUNC program is used to repeat a chosen sample value out to the end.
5. Read back the truncated undegraded pulse and create the compensation function using the DECONVOLVE command. The example below assumes the truncated pulse is already in array A.

```
DRE,B,TUNDG20
DEC,A,B,C           - compensation in C
PLOT,C              - see Fig. 11
```

The compensation function should not have values more than about 10% of maximum outside the first division. A typical compensation function is shown in Fig. 11.

6. The quality of the compensation function is checked by first restoring the original degraded pulse.

```
DRE,A,DGRP20        - get the degraded pulse
CON,A,C,B           - store restored pulse in B
```

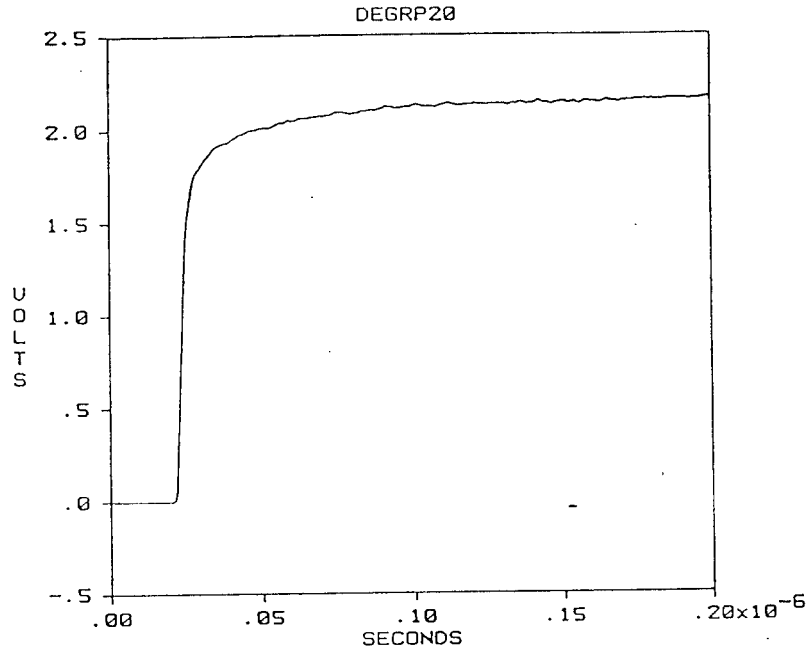


Figure 9. Cable degraded pulse from PBFA II water section.

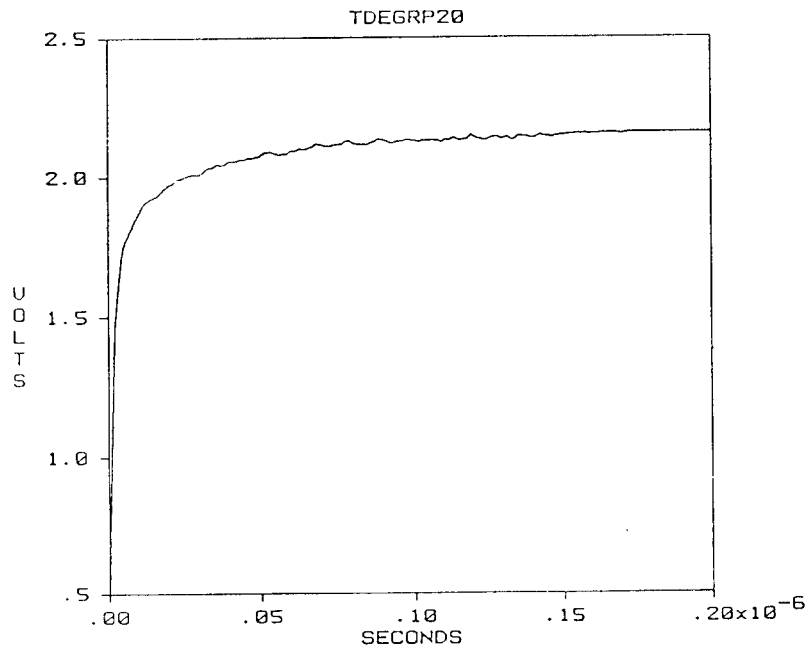


Figure 10. Cable degraded pulse truncated at 25.% of top amplitude.

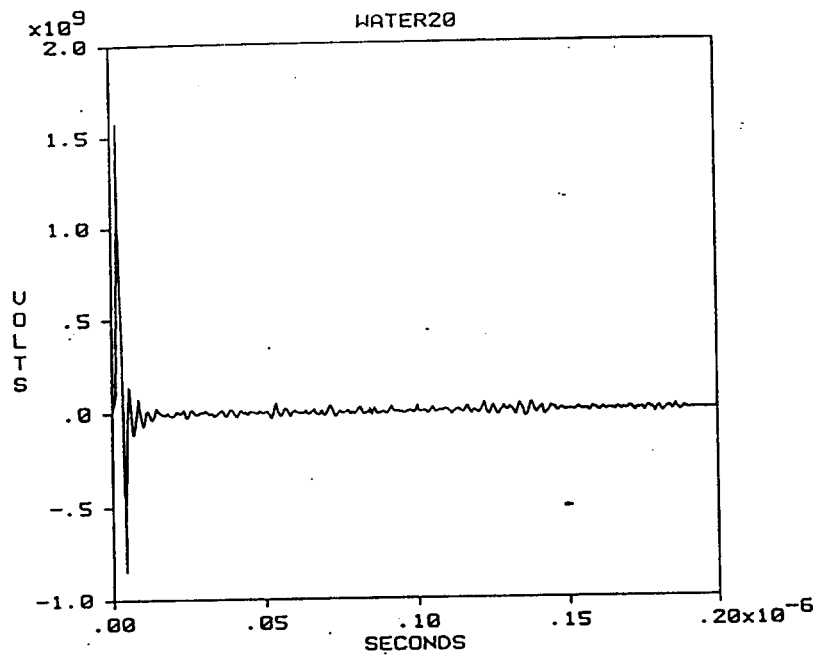


Figure 11. Compensation impulse response function computed from undegraded and degraded pulses in Figs. 8 and 10.

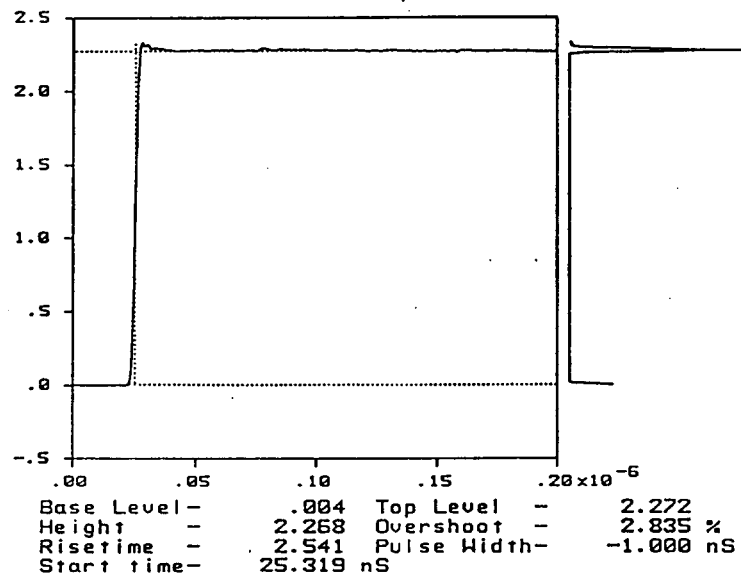


Figure 12. Compensated pulse analyzed for risetime, overshoot, and amplitude. The histogram of pulse amplitudes is shown to right of pulse.

The most objective method for checking the quality of the compensation on the restored signal is to use the IDR command ANALYZE. This command will compute the risetime and overshoot of the restored pulse. It will plot the pulse, a histogram of the amplitude levels and print the measurement results. An example command is

ANA,B

For 7912AD sweep speeds of 20 ns/div or less the compensation function should be adjusted to get the risetime to about 1.9 ns and the overshoot under 3%. An example of a good compensation result in an ANALYZE command format is shown in Fig. 12. Generally lower truncation levels will produce faster risetimes but higher overshoots. If these two conditions cannot be met simultaneously, a different 7912 should be used. For sweep speeds of 50 ns/div and greater only the overshoot parameter is valid. In computing compensation functions, the operator should start with too high of an overshoot; then the truncation level should be raised until the overshoot is less than 3%.

7. When an acceptable restored pulse has been achieved, the compensation function should have its time shift accounted for, and then be labeled and saved.

XFR C F	- save original
CHA F 3 15	- change number of points for an expanded plot
PLO F	
CUR	- locate the peak
X= .5678E-8	- time to peak
Y= .2345E-9	
CHA C 5 -.5678E-8	- change array tzero
LAB,C,WATER20	- water sec. 20ns
SOP,CABCF	- open compensation file
SWR,C	- save compensation function

Otherwise repeat steps 5 and 6. The user should try at least three different truncation levels for each cable to get a feel for how the risetime and overshoot vary.

This completes the time-domain method for computing cable compensation functions. This process must be repeated for each sweep speed and for each different cable run.

As noted previously, acceptable cable compensation functions for 6880 may be computed using this method if multiple undegraded and degraded pulses are averaged to eliminate noise and differences in step response among the CCD paths.

3.2 Analytical Frequency Domain Compensation Method

This section describes in detail how to compute analytical cable compensation functions. Again the interactive data reduction program IDR is used. Although measured degraded and undegraded pulses are not required to compute the compensation filter using the analytical method, they are required to test the results. As described in the preceeding section, a shot data file should be created using AQD for storing all waveforms generated for testing the compensation function. The IDR command for computing the compensator is

CABLE COMPUTE

The program will prompt the user for the various parameters with the message:

Enter Cable 3 dB Freq (MHz), Comp 3 dB Freq, Delta T (ns) number of points, rolloff exp, delay and cable exp

The compensator will be stored in array C
The cable impulse response will be stored in D.

An example reply is

37. 350. .742 1024 4 15. .56

The first parameter is the 3 dB cutoff frequency of the cable run in MHz. This value may either be measured or determined by running the cable attenuation program CAT. The second parameter is the desired 3 dB frequency of the cable compensation filter in MHz. The third parameter is the time step between samples in the filter in nanoseconds. This value should be the same as the time step in the data to be compensated. The fourth parameter is the number of points in the filter. The time step and the number of points define the record length in time of the filter. The filter must be long enough so the degraded step response reaches its final value. A value of 1024 is adequate for most cables. The number of points should be a power of 2 because the compensator filtering is done in the frequency domain using FFT's. The fifth parameter is the rolloff coefficient, p , in equation (28).

The sixth parameter is the time delay required to make both the cable and compensator responses causal. Typical values are 10 to 15 ns. The results can be checked by plotting the compensation and cable impulse responses which are returned in arrays C and D respectively. Examples of good and bad delays are shown in Fig. 13. Note that the program automatically accounts for the time delay by storing the delay in the t-zero entry for the array.

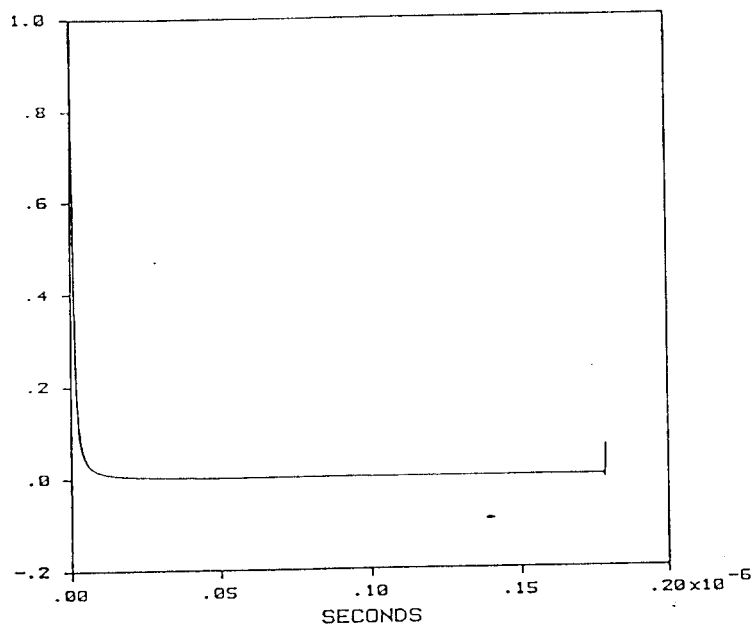


Figure 13a. Cable impulse response computed using analytical method with a delay time of 0 ns.

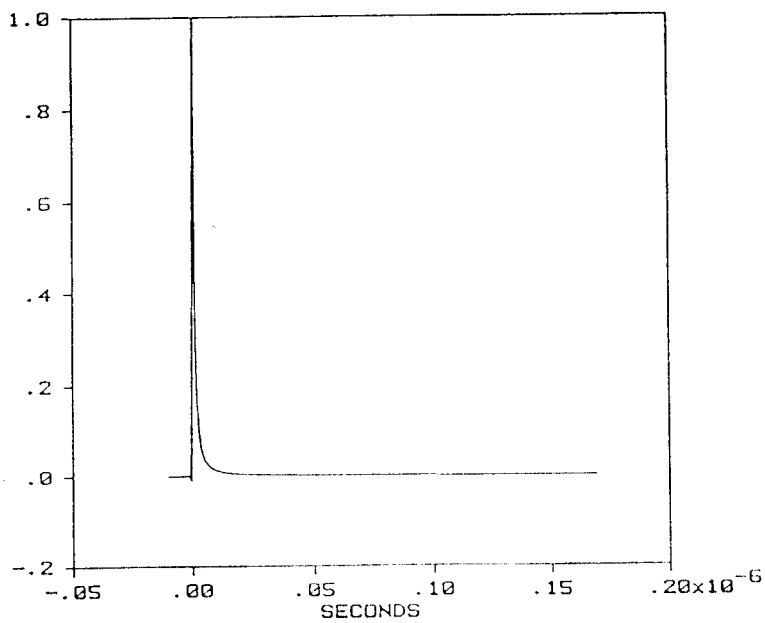


Figure 13b. Cable impulse response computed using a delay of 10 ns.

The last parameter is the cable rolloff rate or the parameter β in equation (29). The theoretical value is .5. However actual values are typically .55 to .57. The values for the cables being used can be obtained by running the cable attenuation program CAT. If a cable run contains multiple different types of cables, an averaged value for the exponent should be used. All the values shown in the example above are typical for LeCroy 6880 recorders. An example compensation function using this method is shown in Fig. 14.

As noted above, undegraded and degraded pulses are required to test the compensation function. For this application the pulses do not have to look like steps, i.e., the pulse can return to zero within the record. Section 3.1 described the method for recording the test pulses using 7912AD's. The commands for recording the pulses using LeCroy 6880 digitizers are shown below. The 6880's do not have an internal trigger. Thus a pulser with an external trigger must be used. Both the Tektronix 506 and the Picosecond Pulse Labs 2500 have this feature.

1. Route the pulse and the trigger to the 6880. The pulse cable should be as short as possible. The pulse voltage should be .35 to .40 volts.

SEL 35	- use unit 35, a 6880
SET INPUT A	- use the A input
SET DELAY -100	- set 100 ns pretrigger
SET OFFSET 200	- positive offset for a PG 506
DIG	- record the pulse. Apply trigger if required.
ACQ,A,0,1000	- read the raw data. The sample time range chosen should include at least 25 ns of baseline.
PLOT A	
CUR X,Y	- locate baseline with horz cursor
SUB A X	- subtract baseline
LAB A UNDGR	- label signal
DWR A	- save in disc file

The steps for recording the degraded pulse are the same except for the signal label. The compensation function should be tested using the CONVOLVE command described in step 6 in section 3.1.

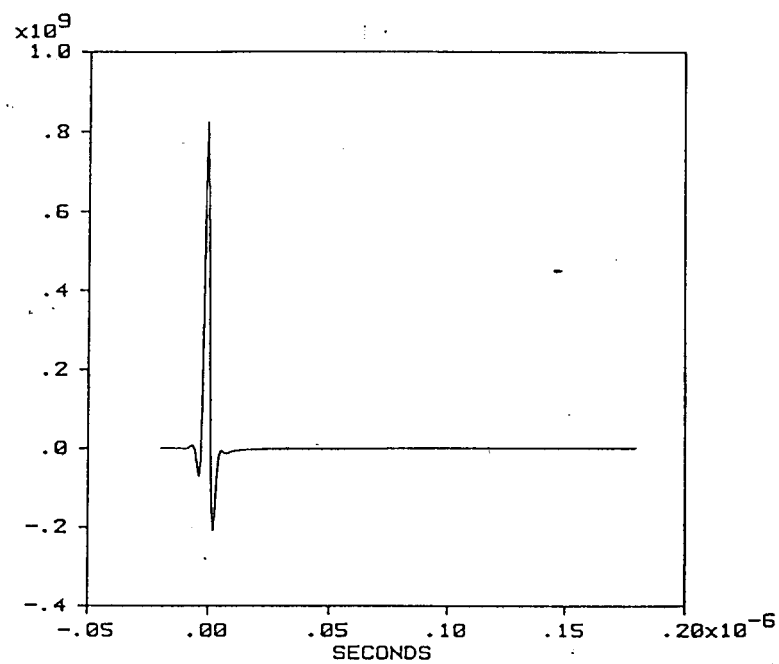


Figure 14. Cable compensation impulse response function computed using analytical method.

3.3 Comparison of Methods

This section compares the two methods of computing cable compensation functions. The results are given for 7912AD's at 20 ns/div for the PBFA II water section diagnostic cables. This run consists of the following cable segments:

50 ft RG-214	- shielded jumper in tank
145 ft RG-331	- fixed cable running from tank to screen room.
25 ft RG-214	- cable from feedthru panel to coax switch
65 ft RG-214	- cable from coax switch to 7912AD

The theoretical 3 dB bandwidth for this cable run is 42.7 MHz. The degraded and undegraded pulses for this run for a 7912AD at 20 ns/div were shown in Figs. 7 and 9. The compensation function corresponding to a toe truncation level of 25% of maximum was shown in Fig. 11. The resulting compensated pulse was shown in Fig. 12. The analytical compensation function was shown in Fig. 14. The parameters used in computing the analytical function were:

Cable 3 dB frequency	- 42.7 MHz
Compensator 3 dB frequency	- 250.0 MHz
Sample interval	-.3906 ns
Number of points	- 512
Compensator rolloff exponent	- 4
Time delay	- 10 ns
Cable rolloff exponent	- .56

Fig. 15 shows the compensated pulse for this function in IDR ANALYZE command format. Fig. 16 shows a comparison of the two methods using the IDR command COMPARE. Fig. 17 shows a comparison of the time-domain compensated pulse to the undegraded pulse. And Fig. 18 shows a comparison of the frequency-domain compensated pulse to the undegraded pulse. All compensations are in excellent agreement.

Figs. 19 and 20 show the frequency responses of the time-domain and frequency-domain compensation functions respectively.

Figs. 21 a,b,c,d show the results of cable compensation on an actual data signal. The signal is a di/dt waveform from an ion B-dot monitor on a PBFA II experiment. Since it is a time-derivative signal, it has very high frequency components. Fig. 21a shows the uncompensated signal as recorded on a 7912AD with a 7A16P plug-in and a sweep speed of 20 ns/div. Fig. 21b shows this signal compensated by a time-domain software cable compensation function. Fig. 21c shows the same signal with a time-domain software compensation function applied to data taken on a LeCroy 6880. The signal peaks in the 7912AD signal are noticeably larger than in the 6880 signal. The problem is that the noise in the 6880 limits the bandwidth achievable in time-domain compensations. Fig 21d shows an overlay plot of the 3 signals.

Figs. 22 a,b,c,d show comparisons of software compensation and hardware equalizer filters on a cable-degraded step pulse from a Picosecond Pulse Labs Model 2500 pulse generator. All the pulses were

recorded on a 7912AD with a 7A29P amplifier at 10 ns/div. This amplifier yields an overall bandwidth of 500 MHz in the 7912AD. Fig. 22a shows the pulse directly into the recorder. Fig. 22b show the result of a time-domain software compensation on the pulse degraded by a cable run similar to the PBFA II water section to 7912AD run described previously. This is a very aggressive compensator, i.e. a low toe-truncation level was used. The aggressiveness of the compensation causes the slight ringing on the top of the pulse. This compensator was designed to maximize the pulse risetime. Figs. 22c and 22d show the results of recording the pulse out of the same cable run and two different Prodyne Services hardware cable equalizers. The output pulses from both hardware equalizers show some undershoot when compared to the input pulse in Fig. 22a. But the pulse from serial #81 is noticeably better than the one from serial #90. Both hardware equalized pulses have a better risetime than the software compensated pulse.

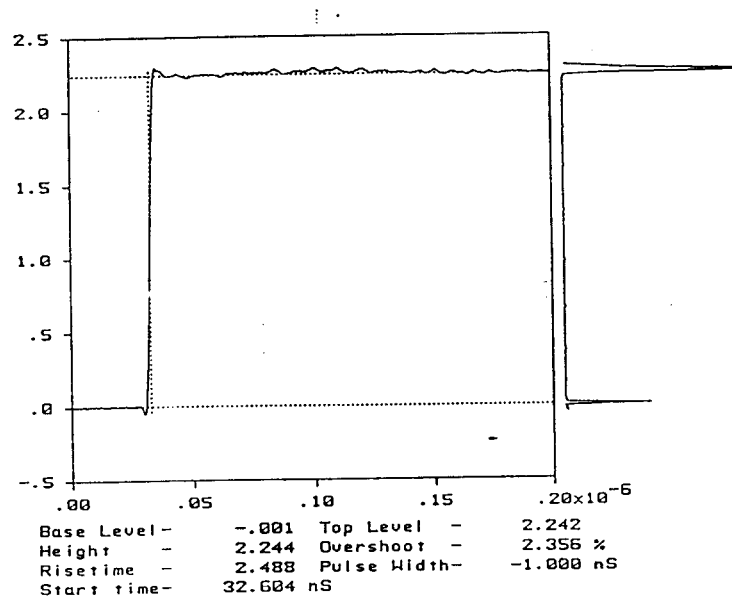


Figure 15. Pulse compensated using analytical compensation function with risetime, overshoot and histogram.

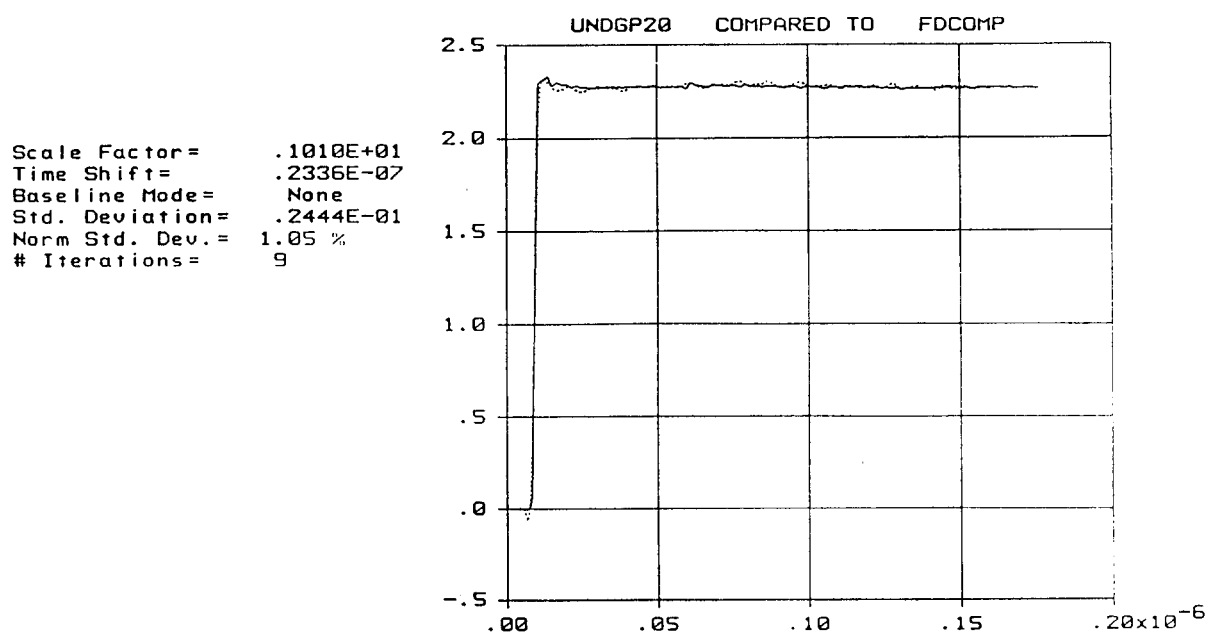


Figure 16. Comparison of time- and frequency-domain compensated pulses.

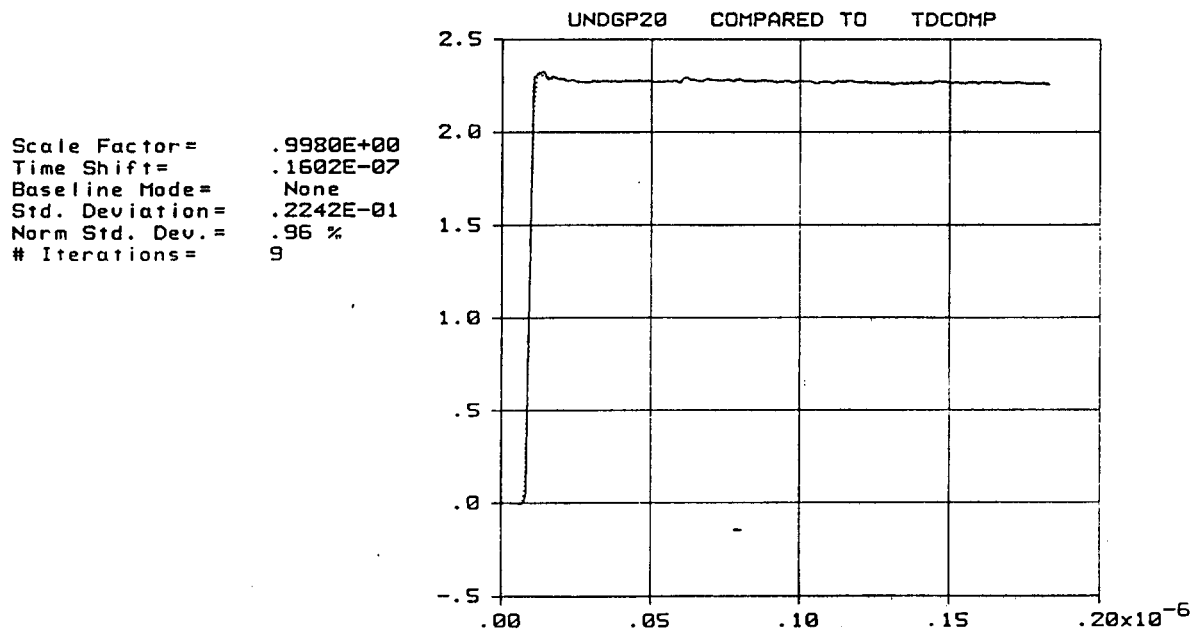


Figure 17. Comparison of time-domain compensated pulse to undegraded pulse.

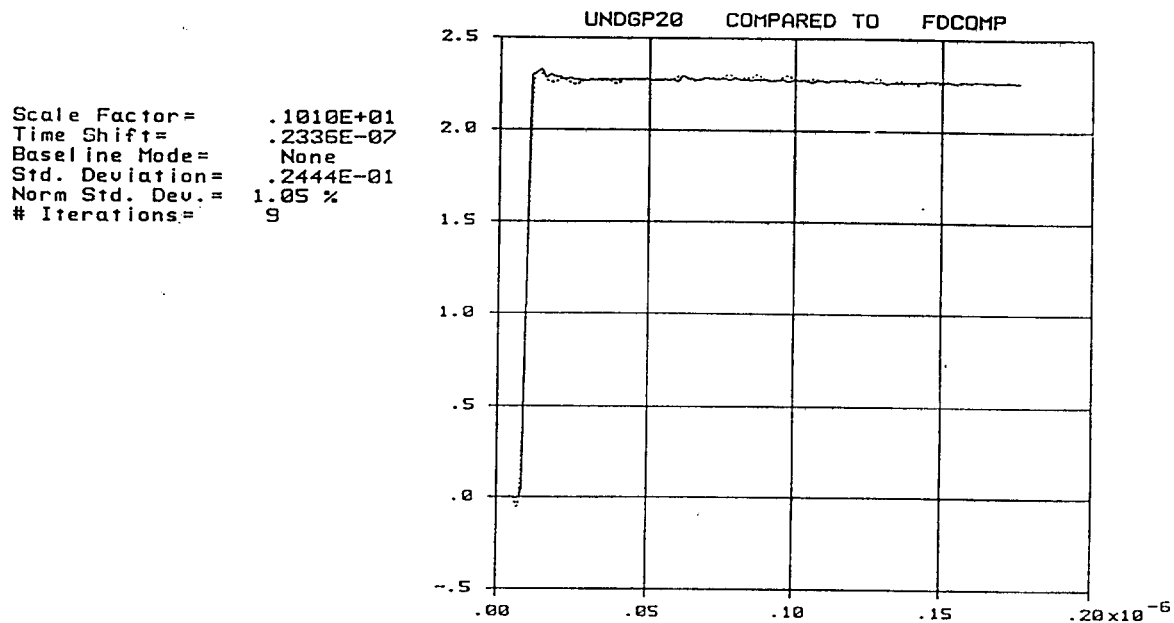


Figure 18. Comparison of frequency-domain compensated pulse to undegraded pulse.

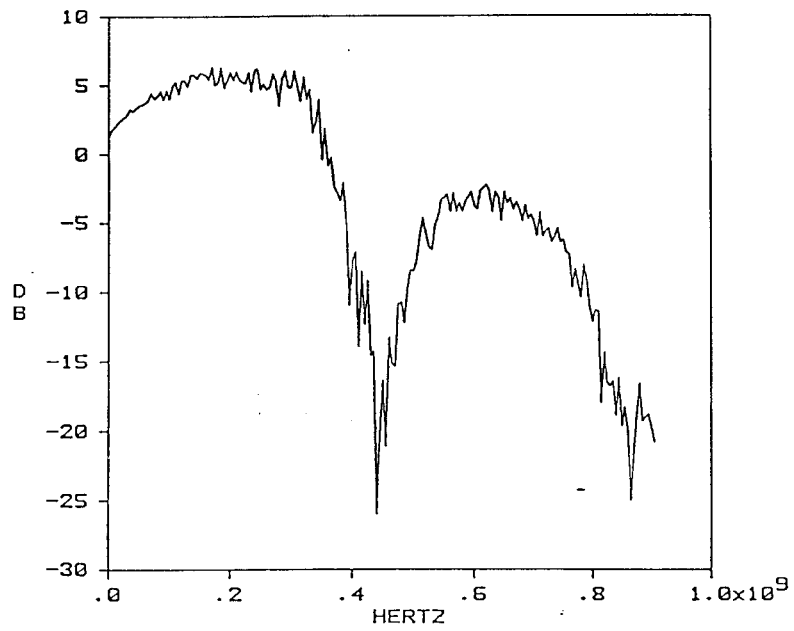


Figure 19. Frequency response of time-domain compensation function.

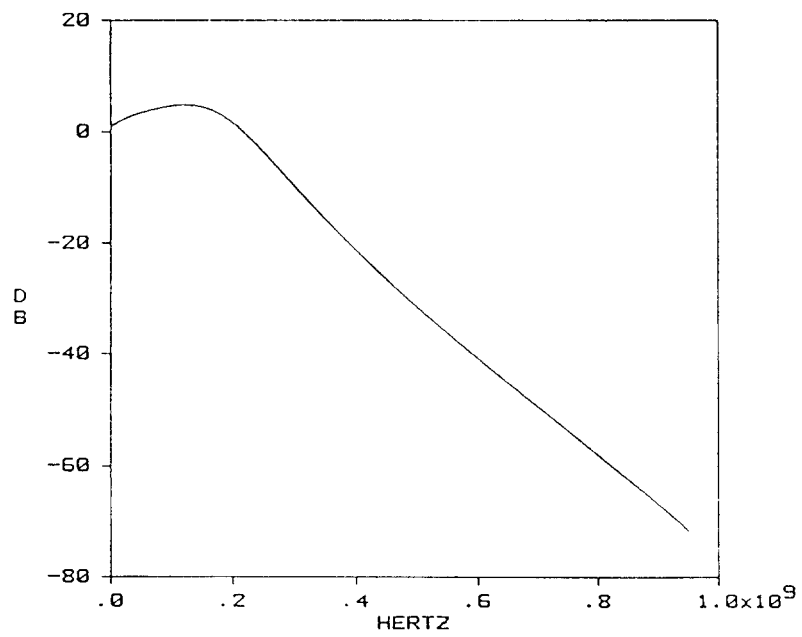


Figure 20. Frequency response of frequency-domain compensation function.

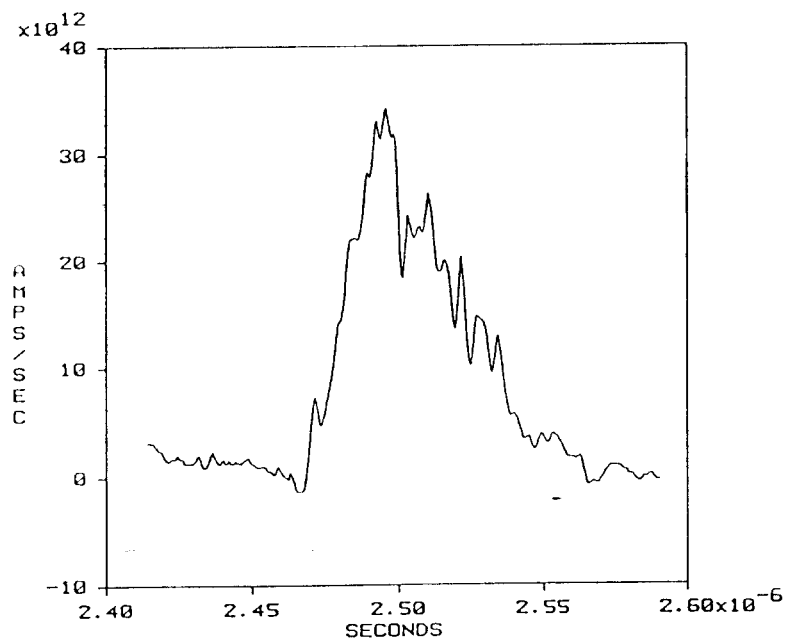


Figure 21a. Uncompensated signal from a PBFA II ion B-dot (di/dt) monitor and recorded on a 7912AD at 20 ns/div.

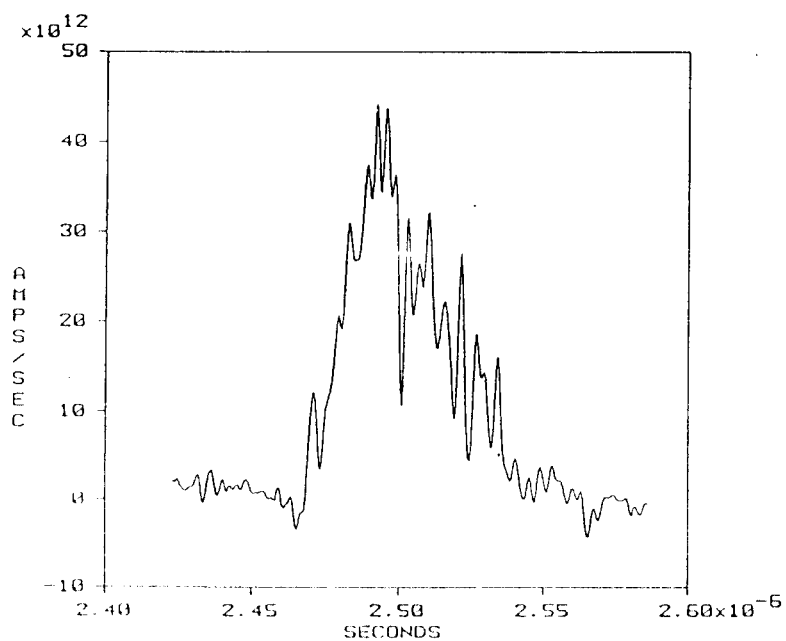


Figure 21b. Ion B-dot signal recorded on a 7912AD and compensated using a time-domain software compensation filter.

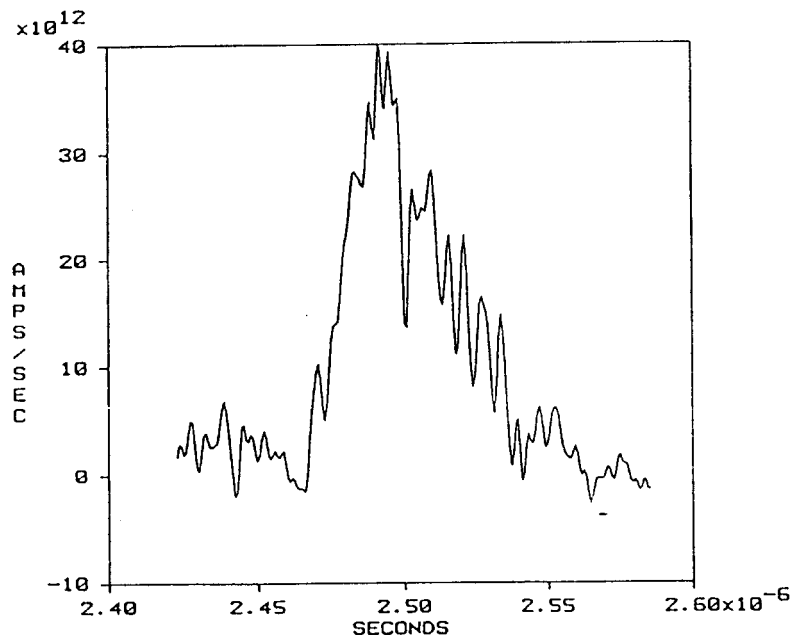


Figure 21c. Ion B-dot signal recorded on a 6880 and compensated using a time-domain compensation filter.

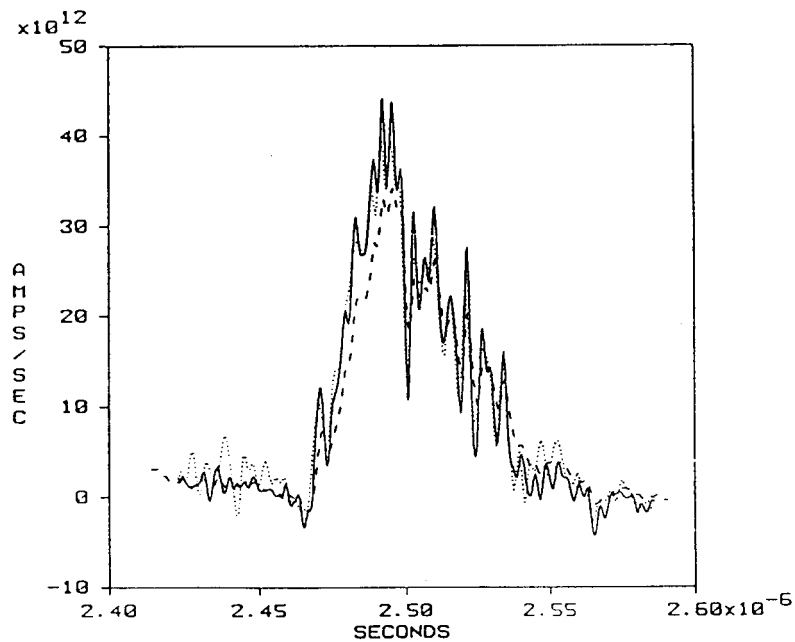


Figure 21d. Overlay plot of the signals in Figs. 21 a,b,c. Solid line is compensated 7912AD signal. Dotted line is compensated 6880 signal. Dashed line is uncompensated 7912AD signal.

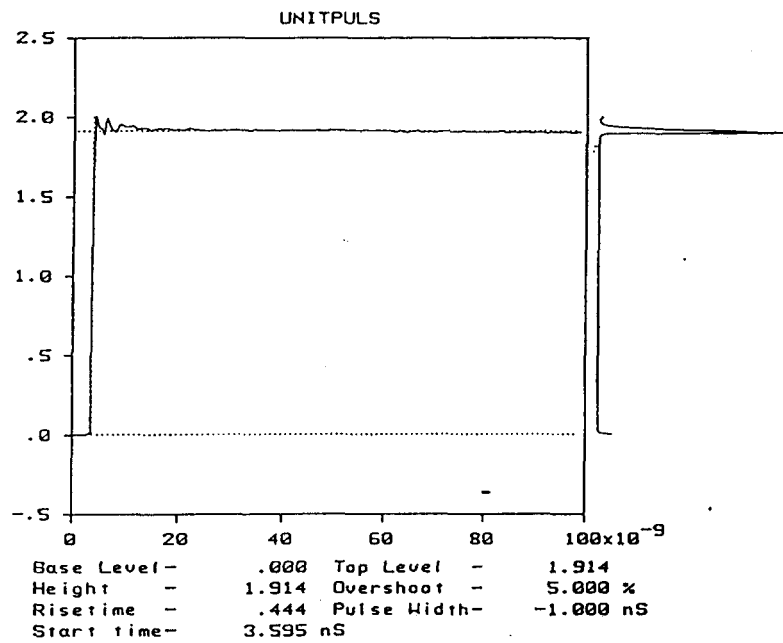


Figure 22a. Step pulse from a Picosecond 2500 pulser recorded directly into a 7912AD with a 7A29P amplifier at 10 ns/div.

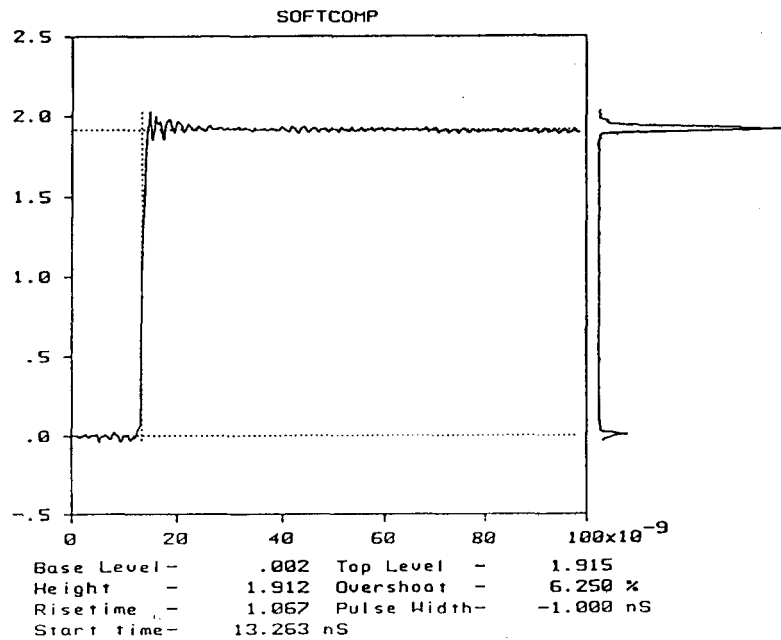


Figure 22b. Cable degraded step pulse restored by a very aggressive time-domain software compensation filter.

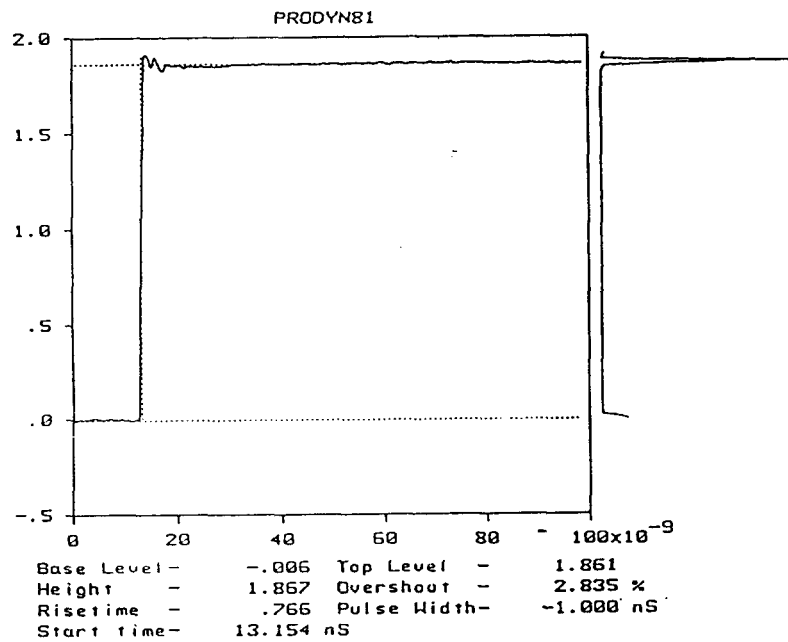


Figure 22c. Cable degraded step pulse restored by a hardware equalizer, serial #81.

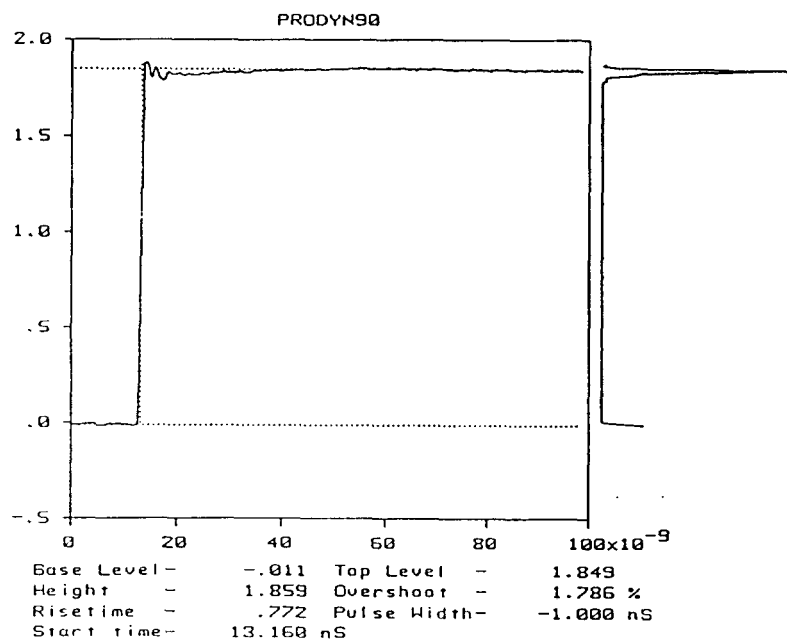


Figure 22d. Cable degraded step pulse restored by a hardware equalizer, serial #90.

3.4 Impulse Test Results

In order to evaluate how software cable compensation performs on narrow pulses, the method was tested with impulse-type pulses. The pulses were generated using an E&H Model 125B pulse generator. The pulses were recorded using a Tektronix 7912AD with a 7A29P vertical amplifier. The cable run tested uses shorter and higher bandwidth cables than the standard PBFA II experiment chamber cable run. The test run has a theoretical 3 dB bandwidth of 81 MHz vs. 48 MHz for the normal run. The pulses tested had nominal pulse widths of 5, 3, and 1 ns. Figures 23 a,b,c show both the pulse recorded straight into the 7912AD and the software compensated pulse recorded at the output of the test cable run. The peak pulse amplitudes for all pulses are given in the table below.

Nominal Pulse Width	Undegraded Pulse Amplitude	Cable compensated Pulse Amplitude
5 ns	10.04 v	9.78
3 ns	10.05 v	9.71
1 ns	7.11 v	5.91

The peak amplitude errors are 3% for the 3 and 5 ns pulses and about 17% for the 1 ns pulse. The reason for the errors can be seen in Fig. 24 a,b. Fig. 24a shows the frequency spectrum of the undegraded 1 ns pulse, solid line, and the cable compensated pulse, dotted line. Note that the two begin to differ near 500 MHz. Fig. 24b shows the frequency response of the cable compensation function used. The compensation function frequency response ceases to rise near 500 MHz. The function would have to continue to rise to compensate for the cable losses at the higher frequencies. The high frequency gain of the cable compensation function was limited by noise in the 7912AD and by the limited signal content in the undegraded step pulse at the higher frequencies.

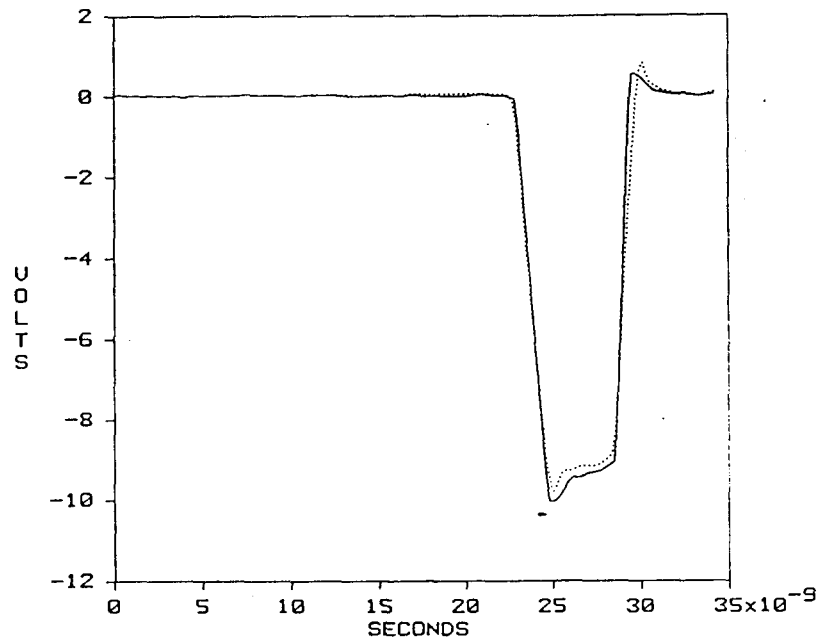


Figure 23a. 5 ns nominal width pulse recorded directly at a 7912AD, solid line, and software compensated pulse recorded at the end of an 81 MHz 3dB bandwidth cable run, dotted line.

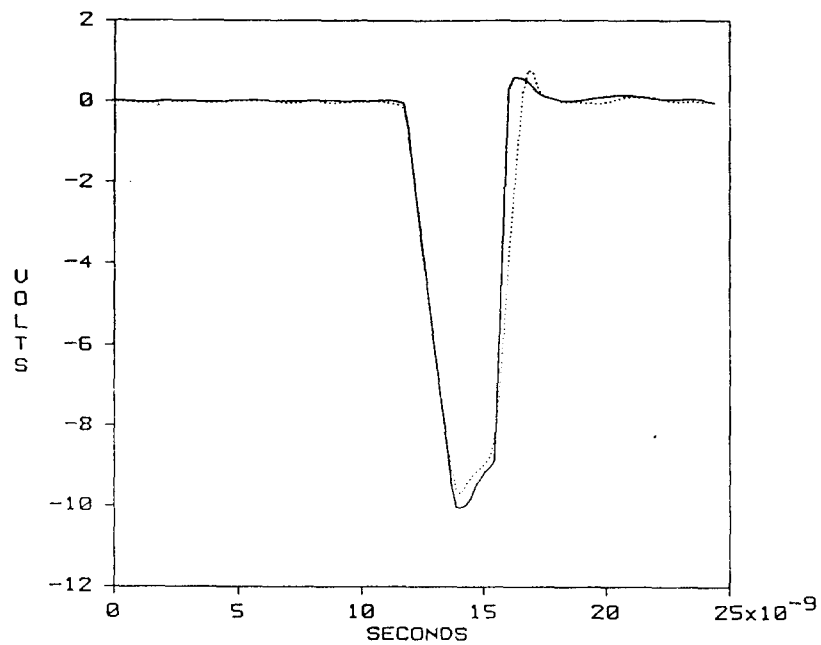


Figure 23b. 3 ns nominal width pulse recorded directly at a 7912AD, solid line, and software compensated pulse recorded at the end of an 81 MHz 3dB bandwidth cable run, dotted line.

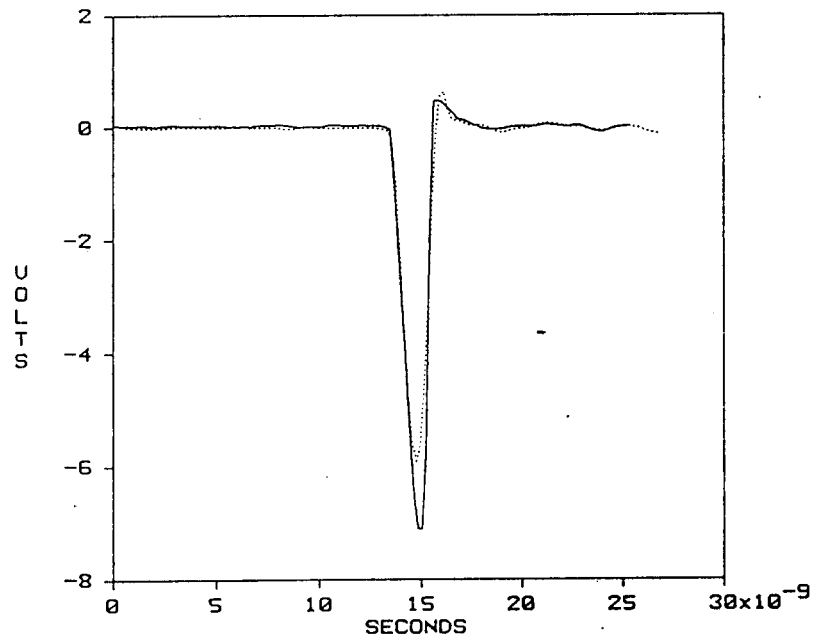


Figure 23c. 1 ns nominal width pulse recorded directly at a 7912AD, solid line, and software compensated pulse recorded at the end of an 81 MHz 3dB bandwidth cable run, dotted line.

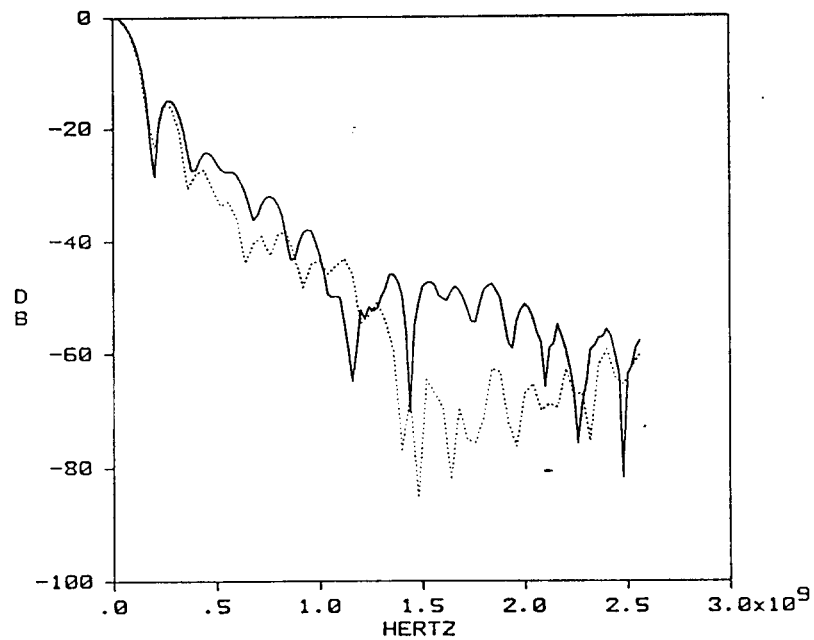


Figure 24a. Frequency spectrum of the undegraded 1ns pulse, solid line, and the cable compensated pulse, dotted line.

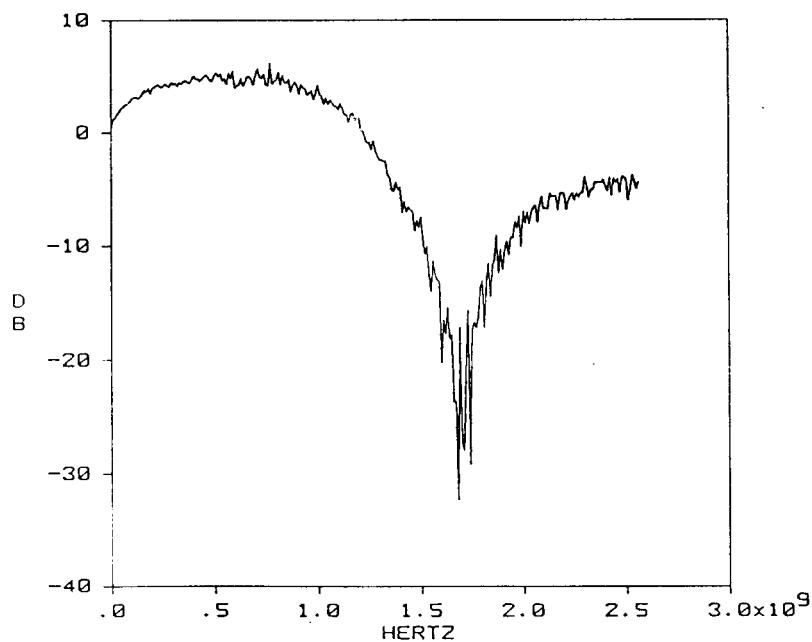


Figure 24b. Frequency response of the cable compensation function used in impulse tests.

4. Conclusion

This paper presents two techniques for computing cable compensation functions, a time domain method using applied step pulses and an analytical frequency domain method. The time domain method has been successfully used in various pulsed power data acquisition systems with Tektronix 7912 Transient Digitizers and more recently with LeCroy 6880s. The analytical method has not been tested thoroughly over a full range of cable runs and recorder noise levels yet. The signal restorations using software cable compensation were shown to be almost as good as those using hardware equalizers.

The high frequency response of software cable compensation is normally limited by noise generated by the waveform recorder rather than by EMP cable pickup noise. At high frequencies, the cable compensation is primarily amplifying noise. Hardware equalizers do not have this limitation since they boost the signal frequency response before the signal enters the recorder.

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9133 W. B. Boyer (25)

3141 S. A. Landenberger (5)
3151 W. L. Garner (3)
3154-1 C. H. Dalin (8) for DOE/OSTI
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